REPORT DOCUMENTATION PAGE

Form Approved
OMB No. 0704-0188

1. REPORT DATE *(DD-MM-YYYY)*	**2. REPORT TYPE**	**3. DATES COVERED** *(From - To)*
JUNE 2013	JOURNAL ARTICLE (Post Print)	JUN 2010 – FEB 2011

4. TITLE AND SUBTITLE

COGNITIVE CODE-DIVISION LINKS WITH BLIND PRIMARY-SYSTEM IDENTIFICATION

5a. CONTRACT NUMBER IN HOUSE

5b. GRANT NUMBER FA8750-11-1-0016

5c. PROGRAM ELEMENT NUMBER 62702F

6. AUTHOR(S)

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5d. PROJECT NUMBER ANCL

5e. TASK NUMBER 62

5f. WORK UNIT NUMBER 07

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N/A

9. SPONSORING/MONITORING AGENCY NAME(S) AND ADDRESS(ES)

Air Force Research Laboratory/Information Directorate
Rome Research Site/RITE
525 Brooks Road
Rome NY 13441-4505

10. SPONSOR/MONITOR'S ACRONYM(S)
AFRL/RI

11. SPONSORING/MONITORING AGENCY REPORT NUMBER
AFRL-RI-RS-TP-2013-030

12. DISTRIBUTION AVAILABILITY STATEMENT
APPROVED FOR PUBLIC RELEASE; DISTRIBUTION UNLIMITED. PA Case Number: 88ABW-2010-4868
DATE CLEARED: 08 SEP 2010

13. SUPPLEMENTARY NOTES

14. ABSTRACT
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15. SUBJECT TERMS
Channel capacity , Cognitive radio , Interference , Multiaccess communication , Receivers , Reliability , Signal to noise ratio

16. SECURITY CLASSIFICATION OF:			**17. LIMITATION OF ABSTRACT**	**18. NUMBER OF PAGES**	**19a. NAME OF RESPONSIBLE PERSON**
a. REPORT	**b. ABSTRACT**	**c. THIS PAGE**			MICHAEL J. MEDLEY
U	U	U	UU	12	**19b. TELEPHONE NUMBER** (*Include area code*) N/A

Cognitive Code-Division Links with Blind Primary-System Identification

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***Abstract*—We consider the problem of cognitive code-division channelization (simultaneous power and code-channel allocation) for secondary transmission links co-existing with an unknown primary code-division multiple-access (CDMA) system. We first develop a blind primary-user identification scheme to detect the binary code sequences (signatures) utilized by primary users. To create a secondary link we propose two alternative procedures –one of moderate and one of low computational complexity– that optimize the secondary transmitting power and binary code-channel assignment in accordance with the detected primary code channels to avoid "harmful" interference. At the same time, the optimization procedures guarantee that the signal-to-interference-plus-noise ratio (SINR) at the output of the maximum SINR linear secondary receiver is no less than a certain threshold to meet secondary transmission quality of service (QoS) requirements. The extension of the channelization problem to multiple secondary links is also investigated. Simulation studies presented herein illustrate the theoretical developments.**

***Index Terms*—Blind user identification, code-channel allocation, code-division multiple-access, cognitive radio, dynamic spectrum access, power allocation, signal-to-interference-plus-noise ratio.**

I. Introduction

With the rapid proliferation of a variety of consumer oriented wireless devices, demand for access to radio spectrum has been growing dramatically and the limited available spectrum is becoming increasingly congested. At the same time, location-dependent bands of pre-licensed radio spectrum may experience low utilization [1]. Cognitive radio (CR) is an emerging technology aiming at improving spectrum utilization efficiency by allowing secondary users/networks to opportunistically share radio spectrum originally licensed by primary users/networks without causing "harmful" interference to them [2]-[5].

Cognitive radio networks can be categorized according to two modes of operation: *cooperation* mode and *coexistence* mode [6]. In cooperation mode, primary users cooperate with secondary users and share information to avoid mutual interference. In coexistence mode, there is no form of cooperation and secondary users must have the ability to detect the presence of primary users [7], [8] and change behavior accordingly to avoid mutual interference.

Past work in the young field of cognitive code-division channelization includes coexistence power control [9] as well as distributed resource allocation of spectral bands, power, and data rates among multiple secondary users for multi-carrier CDMA systems [10]. Cooperation-mode bit rate and spreading factor adjustments for a secondary CDMA system under interference-minimizing code assignments were carried out in [11]. In [12], [13], a secondary maximum signal-to-interference-plus-noise ratio (SINR) code-division link is designed subject to SINR requirements for the primary system which is presumed known (cooperation-mode cognitive radio). Outside the framework of cognitive code-division altogether, interesting work in the form of joint beamforming and power allocation algorithms was reported in [14]-[18]. In particular, in [14] the radio frequency spectrum of interest was divided into a set of multiple orthogonal channels and was shared between primary and secondary networks using orthogonal frequency division multiple access (OFDMA). In [15]-[17], joint spatial-channel and power allocation algorithms for cognitive radio networks were developed. In [18], the authors provide a solution for leasing spectrum for a fraction of time to secondary users based on the idea that secondary nodes can earn spectrum access in exchange for transmission assistance to the primary link (cooperative communications paradigm).

In this work, we focus on coexistence-mode cognitive radio and investigate the problem of establishing a secondary code-division link coexisting with an *unknown* primary CDMA system. In particular, we investigate –to the best of our knowledge for the first time in the context of cognitive radio– the problem of blindly identifying the binary codes/signatures utilized by primary users when neither channel state information nor pilot signaling (training sequence) is available. Then, we study the design of a power and binary-code-channel allocation protocol for the secondary link that will not cause "harmful" interference to the existing primary users. Since post-processing interference sensing is not feasible in coexistence mode cognitive radio, to quantify "harm" we use the periodic-total-square-correlation (PTSC) interference metric in our optimization problem as the mathematical means to protect co-channel primary users [2], [19], [20]. At the same time, to satisfy quality-of-service (QoS) requirements for the secondary link, the power and code-division optimization

Manuscript received July 13, 2010; revised March 23, 2011 and August 26, 2011; accepted September 1, 2011. The associate editor coordinating the review of this paper and approving it for publication was D. Dardari.

This work was supported by the U.S. Air Force Research Laboratory under Grants FA8750-08-1-0063 and FA8750-11-1-0016. Approved for Public Release; Distribution Unlimited: 88ABW-2010-4868.

This paper was presented in part at IEEE MILCOM, San Jose, CA, November 2010.

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Digital Object Identifier 10.1109/TWC.2011.091911.101260

problem is constrained to have SINR at the output of the maximum SINR linear receiver of the secondary link no less than a certain threshold. We recognize that the above described fundamental cognitive code-division radio formulation is, regretfully, a non-convex NP-hard problem. Yet, using herein existing SINR-maximization signature design methodologies we are able to develop novel, realizable suboptimum solutions of varying computational complexity with excellent cognitive system performance characteristics as demonstrated by simulation studies included in this paper. The theoretical developments and experiments can be readily extended to cover multiple secondary links alongside the primary CDMA system.

The rest of this paper is organized as follows. The code-division cognitive radio problem formulation is presented in Section II. A novel primary user identification algorithm is introduced in Section III. The power allocation and signature design procedure for a secondary link is developed in Section IV. In Section V, we extend the procedure to solve for multiple secondary links. Simulation results are presented in Section VI and, finally, a few conclusions are drawn in Section VII.

II. System Model and Problem Formulation

The following notation is used throughout this paper. Boldface lower-case letters indicate column vectors and boldface upper-case letters indicate matrices; $\mathbb{C}$ denotes the set of all complex numbers, $()^T$ and $()^H$ denote the transpose and transpose-conjugate operation, respectively; $\mathbf{I}_L$ is the $L \times L$ identity matrix, $\mathfrak{Re}\{\cdot\}$ denotes the real part of a complex number, $\operatorname{sgn}\{\cdot\}$ denotes zero-threshold quantization, and $\mathbb{E}\{\cdot\}$ represents statistical expectation. Finally, $|\cdot|$, $\|\cdot\|$, and $\|\cdot\|_F$ are the scalar magnitude, vector norm, and matrix Frobenius norm, respectively.

In the following, we consider a primary code-division system with a primary transmitter PT and K primary receivers $PR_i, i = 1, 2, \ldots, K$, as shown in Fig. 1. The primary transmitter (for example, base station) PT communicates downlink with the K primary receivers $PR_i, i = 1, 2, \ldots, K$, over distinct code-division channels defined by individual normalized binary codes/signatures $\mathbf{s}_i = \frac{1}{\sqrt{L}}\{\pm 1\}^L$, $i = 1, 2, \ldots, K$, where L is the signature length (system processing gain). We consider also a potential concurrent secondary code-division link in the spectrum band of the primary downlink channel between a secondary transmitter ST and receiver SR. If the primary system is frequency-division-duplex (FDD), then the secondary ST-to-SR link will work on the frequency licensed to the downlink channel of the primary system and can operate at any time. If the primary system is time-division-duplex (TDD), then the secondary link can transmit data only during the downlink slot of the primary system. The secondary communication link is activated, whenever possible, with a (normalized) binary signature $\mathbf{c} = \frac{1}{\sqrt{L}}\{\pm 1\}^L$ and transmitting power $P > 0$. All transmitted signals, primary and secondary when appropriate, are assumed/modeled to propagate over multipath Rayleigh fading channels and experience additive white Gaussian noise (AWGN).

We first assume that the secondary transmitter ST is quiet and we examine how the signal sent by PT is observed by SR. After carrier demodulation, chip matched filtering and

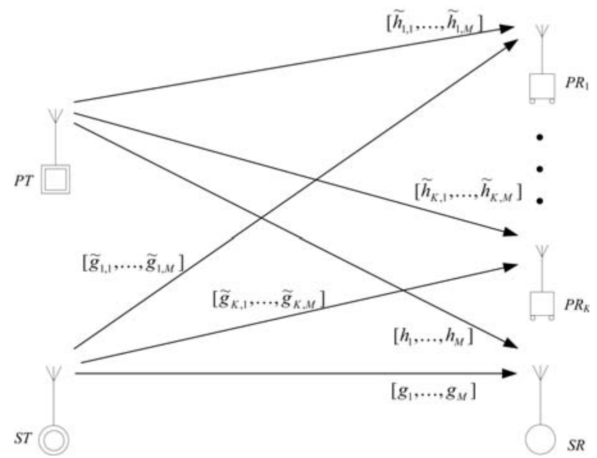

Fig. 1. Primary/secondary code-division system model of a primary transmitter PT, K primary receivers $PR_i, i = 1, 2, \ldots, K$, and a secondary transmitter/receiver pair ST, SR (all received signals exhibit multipath Rayleigh fading).

sampling at the chip rate over a presumed multipath extended data bit period of $L_M = L + M - 1$ chips where M is the number of resolvable multipaths, the observed data vector $\mathbf{y}(n) \in \mathbb{C}^{L_M}$ by SR takes the following general form[1]

$$\mathbf{y}(n) = \sum_{i=1}^{K} \sqrt{E_i} b_i(n) \mathbf{H}\mathbf{s}_i + \mathbf{i} + \mathbf{n}, \quad n = 1, 2, \ldots, \tag{1}$$

where $\mathbf{H} \in \mathbb{C}^{L_M \times L}$ is the multipath channel matrix between PT and SR

$$\mathbf{H} \triangleq \begin{bmatrix} h_1 & 0 & \ldots & 0 & 0 \\ h_2 & h_1 & \ldots & 0 & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ h_M & h_{M-1} & & 0 & 0 \\ 0 & h_M & & 0 & 0 \\ \vdots & \vdots & & \vdots & \vdots \\ 0 & 0 & \ldots & h_M & h_{M-1} \\ 0 & 0 & \ldots & 0 & h_M \end{bmatrix} \tag{2}$$

with entries $h_m \in \mathbb{C}$, $m = 1, \ldots, M$, considered as complex Gaussian random variables to model fading phenomena, $\sqrt{E_i} > 0$ and $b_i(n) \in \{\pm 1\}$ are the amplitude level and nth transmitted bit of primary user i, $i = 1, \ldots, K$, respectively, $\mathbf{i} \in \mathbb{C}^{L_M}$ denotes multipath induced inter-symbol-interference (ISI), and $\mathbf{n}$ is a zero-mean additive white Gaussian noise (AWGN) vector with autocorrelation matrix $\sigma^2 \mathbf{I}_{L_M}$. The information bits $b_i(n)$ are viewed as binary equiprobable random variables that are independent within a user stream (in $n = 1, 2, \ldots$) and across users (in $i = 1, 2, \ldots, K$). Since the effect of ISI is, arguably, negligible for most applications of practical interest where the number of resolvable multipaths is much less than the processing gain, for mathematical convenience we will not consider the ISI terms in our theoretical developments that follow[2]. Thus, the primary users' signal

[1] Equation (1) assumes that SR is symbol-synchronous to PT. While this is not a technical requirement at all for the secondary link design problem set and solved herein based on the periodic cross-channel total squared correlation, it greatly simplifies the presentation and notation of this material.

[2] However, naturally ISI will be considered and accounted for in our simulation studies.

observed by SR in (1) is simplified/approximated by

$$\mathbf{y}(n) = \sum_{i=1}^{K} \sqrt{E_i} b_i(n) \mathbf{H} \mathbf{s}_i + \mathbf{n}, \quad n = 1, 2, \ldots. \tag{3}$$

In our cognitive system model, the secondary link is taken to be chip-synchronous to the primary network (worst case interference scenario) with the same chip rate and symbol-synchronous (see Footnote 1). Also, without loss of generality and for simplicity in notation, we assume that the multipath channels between PT and PR_i, $i = 1, \ldots, K$, PT and SR, and ST and SR, all have the same number of resolvable paths. Then, when the secondary communication link is activated with a (normalized) binary signature code $\mathbf{c} = \frac{1}{\sqrt{L}}\{\pm 1\}^L$ and transmit power $P > 0$, the aggregate signal vector received by SR can be expressed as

$$\mathbf{r}(n) = \sqrt{P} b(n) \mathbf{G} \mathbf{c} + \mathbf{y}(n), \quad n = 1, 2, \ldots, \tag{4}$$

where $\mathbf{G} \in \mathbb{C}^{L_M \times L}$ is the ST to SR channel matrix with multipath channel coefficients $g_m \in \mathbb{C}$, $m = 1, \ldots, M$, and $\mathbf{y}(n)$ is given by (3).

Information bit detection at SR is carried out via linear maximum SINR filtering (or, equivalently, minimum mean square error filtering) as follows

$$\widehat{b}(n) = \operatorname{sgn}\left\{\Re\mathfrak{e}\{\mathbf{w}_{maxSINR}^H \mathbf{r}(n)\}\right\}, \quad n = 1, 2, \ldots, \tag{5}$$

where $\mathbf{w}_{maxSINR} = c\mathbf{R}^{-1}\mathbf{G}\mathbf{c} \in \mathbb{C}^{L_M}$, $c > 0$, is the maximum SINR filter and $\mathbf{R} = \mathrm{E}\{\mathbf{y}\mathbf{y}^H\}$ is the autocorrelation matrix of the $\mathbf{y}(n)$ signal in (3) that constitutes primary-system disturbance to SR. Practically, $\mathbf{R}$ is estimated by averaging over $N \geq L_M$ observation samples $\mathbf{r}(n)$ when ST is silent ($P = 0$), $\widehat{\mathbf{R}}(N) = \frac{1}{N}\sum_{n=1}^{N} \mathbf{r}(n)\mathbf{r}(n)^H = \frac{1}{N}\sum_{n=1}^{N} \mathbf{y}(n)\mathbf{y}(n)^H$. The output SINR of the filter $\mathbf{w}_{maxSINR}$ can be calculated to be

$$\Gamma \triangleq \frac{\mathbb{E}\{|\mathbf{w}_{maxSINR}^H(\sqrt{P} b \mathbf{G}\mathbf{c})|^2\}}{\mathbb{E}\left\{|\mathbf{w}_{maxSINR}^H \mathbf{y}|^2\right\}} = P\mathbf{c}^T\mathbf{G}^H\mathbf{R}^{-1}\mathbf{G}\mathbf{c}. \tag{6}$$

To attain a certain QoS level for the secondary link, we need to jointly design the binary signature $\mathbf{c}$ and the transmitting power P to have the SINR value Γ at the output of the maximum SINR filter $\mathbf{w}_{maxSINR}$ no less than a given threshold $\gamma > 0$, i.e. $\Gamma \geq \gamma$.

At the same time, due to the coexistence of the secondary link with the primary network, interference is introduced to the primary receivers. The secondary link is to be allowed to activate *only if* the interference to each primary receiver is not "harmful." The difficulty is that in cognitive radio networks operating in coexistence-mode, the primary network does not cooperate/talk to secondary users and the latter do not have global knowledge of system parameters, such as the multipath channel coefficients $[\widetilde{h}_{i,1}, \ldots, \widetilde{h}_{i,M}]$ between PT and PR_i, or the channel coefficients $[\widetilde{g}_{i,1}, \ldots, \widetilde{g}_{i,M}]$ between ST and PRi, $i = 1, 2, \ldots, K$ (see Fig. 1), or the primary binary signatures $\mathbf{s}_i$, $i = 1, \ldots, K$, in (3) and the filters utilized by each primary user receiver[3]. Therefore, post-processing interference sensing is not feasible.

[3]In fact, even if global system parameter values (code channels, transmission power values, channel coefficients, receive filters) were to be provided to the secondary system by the primary system, SINR optimized secondary link design under primary-system SINR constraints is a non-convex, NP-hard problem [12], [13].

Toward a realistic, realizable solution to the problem of coexistence secondary link design, we recall that the periodic (cyclic) total squared cross-correlation (PTSC) value [19], [20] is a useful measure to evaluate multiple access interference (MAI) when channels exhibit multipath behavior. In this spirit, we propose to use the PTSC value as a metric to evaluate the interference caused by the secondary link. For notational simplicity, let $\mathbf{s}_{i|l}$, $i = 1, \ldots, K$, denote the *cyclic right-shifted* version of $\mathbf{s}_i \in \frac{1}{\sqrt{L}}\{\pm 1\}^L$, by l bit positions when $l = 0, 1, \ldots, L-1$, and *cyclic left-shifted* version of $\mathbf{s}_i$ by l bit positions when $l = 0, -1, \ldots, 1-L$, (hence, $\mathbf{s}_{i|0} = \mathbf{s}_i$). The PTSC between signatures $\mathbf{c}$ and $\mathbf{s}_i$ for multipath shifts up to lag M is defined as

$$\mathrm{PTSC}(\mathbf{c}, \mathbf{s}_i) \triangleq \sum_{l=-M}^{M} |\mathbf{c}^T \mathbf{s}_{i|l}|^2, i = 1, \ldots, K. \tag{7}$$

In this context, we define the *generalized correlation interference* by a secondary link with power $P > 0$ to the ith primary receiver as

$$\mathcal{I}_i \triangleq P \cdot \mathrm{PTSC}(\mathbf{c}, \mathbf{s}_i), \quad i = 1, \ldots, K. \tag{8}$$

We understand that $\mathcal{I}_i$ serves as a simple "worst-case" measure of the effect of the secondary link on the ith primary user. Then, in this context, the secondary link can be activated by assigning a signature $\mathbf{c}$ and power $P > 0$ if the interference to *every* primary user is less than a threshold $\mathcal{I}_{\mathrm{th}} > 0$: $\mathcal{I}_i < \mathcal{I}_{\mathrm{th}}$ $\forall i = 1, \ldots, K$. If

$$\mathcal{I}_{max} \triangleq max\{\mathcal{I}_i : i = 1, \ldots, K\} \tag{9}$$

is the strongest generalized interference to primary receivers, the activation condition is equivalent to $\mathcal{I}_{max} < \mathcal{I}_{\mathrm{th}}$.

Our objective is to jointly design the binary signature $\mathbf{c}$ and the transmitting power $P > 0$ for the secondary link to minimize $\mathcal{I}_{max}$ under the constraint that the secondary link achieves its pre-determined SINR requirement γ:

$$\underset{P>0,\ \mathbf{c}\in\frac{1}{\sqrt{L}}\{\pm 1\}^L}{\text{minimize}} \ \mathcal{I}_{max} \triangleq max\{P \cdot \mathrm{PTSC}(\mathbf{c}, \mathbf{s}_i) : i = 1, \ldots, K\} \tag{10}$$

$$\text{s. t. } \Gamma \triangleq P\mathbf{c}^T\mathbf{G}^H\mathbf{R}^{-1}\mathbf{G}\mathbf{c} \geq \gamma. \tag{11}$$

Then, if the resulting minimized $\mathcal{I}_{max}$ is indeed less than $\mathcal{I}_{\mathrm{th}}$, the secondary link can be activated; otherwise, it is kept idle.

The cognitive code-division channelization problem formulated in (10), (11) requires knowledge of all active primary-user binary signatures $\mathbf{s}_i$, $i = 1, \ldots, K$, for the evaluation of the created interference in the form of $\mathrm{PTSC}(\mathbf{c}, \mathbf{s}_i)$, $i = 1, \ldots, K$. However, in our assumed coexistence mode of operation, the secondary network cannot obtain such knowledge directly from primary networks. Therefore, before starting to solve the channelization problem in (10), (11), the secondary network needs to blindly detect *(i)* the number of active primary users K and *(ii)* their binary signatures $\mathbf{s}_i$, $i = 1, \ldots, K$. With respect to item (i) (population size identification problem), we can utilize for example the algorithm developed recently in [21]. Due to space limitations, in this paper we do not deal further with this issue and instead assume that K is correctly identified. With respect to item (ii), in the next section we develop an iterative-least-square (ILS)-based procedure that can blindly detect the primary users' binary signatures $\mathbf{s}_i$, $i = 1, \ldots, K$, from the observed primary signal $\mathbf{y}(n)$.

TABLE I
ITERATIVE LEAST-SQUARES PROCEDURE

1) $d := 0$; initialize $\widehat{\mathbf{B}}^{(0)} \in \{\pm 1\}^{K \times N}$ arbitrarily.
2) $d := d + 1$;
$$\widehat{\mathbf{V}}^{(d)} := \mathbf{Y}(\widehat{\mathbf{B}}^{(d)})^T \left[(\widehat{\mathbf{B}}^{(d)})(\widehat{\mathbf{B}}^{(d)})^T\right]^{-1};$$
$$\widehat{\mathbf{B}}^{(d)} := \mathrm{sgn}\left\{\Re\mathfrak{e}\left[\left[(\widehat{\mathbf{V}}^{(d-1)})^H(\widehat{\mathbf{V}}^{(d-1)})\right]^{-1}(\widehat{\mathbf{V}}^{(d-1)})^H\mathbf{Y}\right]\right\}.$$
3) Repeat Step 2 until $(\widehat{\mathbf{B}}^{(d)}, \widehat{\mathbf{V}}^{(d)}) = (\widehat{\mathbf{B}}^{(d-1)}, \widehat{\mathbf{V}}^{(d-1)})$.

III. PRIMARY-SYSTEM IDENTIFICATION

If we denote the (energy inclusive) channel processed signature by

$$\mathbf{v}_i \triangleq \sqrt{E_i}\mathbf{H}\mathbf{s}_i,\ i = 1, 2, \ldots, K, \tag{12}$$

then the observed signal in (3) can be expressed as

$$\mathbf{y}(n) = \sum_{i=1}^{K} \mathbf{v}_i b_i(n) + \mathbf{n} = \mathbf{V}\mathbf{b}(n) + \mathbf{n},\ \ n = 1, 2, \ldots, \tag{13}$$

where $\mathbf{V}_{L_M \times K} \triangleq [\mathbf{v}_1, \ldots, \mathbf{v}_K]$ is the effective signature matrix and $\mathbf{b}(n) \triangleq [b_1(n), \ldots, b_K(n)]^T$ is the vector of bits for all K users at the nth transmission period. If SR is able to collect N observation vectors $\mathbf{y}(n)$, $n = 1, 2, \ldots, N$, then (13) can be rewritten in matrix form as

$$\mathbf{Y} = \mathbf{V}\mathbf{B} + \mathbf{N} \tag{14}$$

where $\mathbf{Y} \in \mathbb{C}^{L_M \times N}$ is the observation matrix, $\mathbf{B} \triangleq [\mathbf{b}(1), \ldots, \mathbf{b}(N)]$ is the $K \times N$ data matrix that contains the N bits transmitted for each of the K primary users, and $\mathbf{N}$ is an $L_M \times N$ Gaussian noise matrix.

To detect the binary signatures $\mathbf{s}_i$, $i = 1, \ldots, K$, we first estimate the channel processed signature set $\mathbf{V}$ from the observation matrix $\mathbf{Y}$. Our approach begins by formulating the signature set estimation problem as a joint detection and estimation problem with the following least squares (LS) solution

$$\widehat{\mathbf{V}}, \widehat{\mathbf{B}} = \arg \min_{\substack{\mathbf{B} \in \{\pm 1\}^{(K \times N)}, \\ \mathbf{V} \in \mathbb{C}^{L_M \times K}}} \|\mathbf{Y} - \mathbf{V}\mathbf{B}\|_F^2 . \tag{15}$$

The above LS solution is maximum-likelihood optimal as long as $\mathbf{N}$ is white Gaussian. In any case, regretfully, joint detection and estimation by (15) has complexity exponential in NK. We consider this cost unacceptable and attempt to reach a quality approximation of the solution by alternating least squares estimates of $\mathbf{V}$ and $\mathbf{B}$, iteratively, as described below.

A. Iterative Least Squares Procedure

The basic idea behind such an iterative least squares (ILS) solution [22]-[25] is to compute an LS update of one of the unknown (matrix) parameters conditioned on a previously obtained estimate of the other (matrix) parameter and continue on until convergence is observed. The iterative least-squares procedure for the solution of our problem in (15) is presented in Table I. Superscripts in Table I denote the iteration index. Derivation details are presented in the Appendix.

We understand that convergence of the developed iterative least squares procedure to a globally optimal LS-solution of (15) is not guaranteed in general. The quality (least-squares fit) of the end convergence point depends heavily on the initialization point and arbitrary initialization –which at first sight is unavoidable for blind primary system identification– offers little assurance that the iterative scheme will lead us to appropriate, "reliable" (close to minimal least-squares fit) solutions. Re-initialization and re-execution[4] of the ILS procedure is always possible but the challenge is how to assess whether solutions returned by the ILS procedure are reliable or not without any side information or pilot signaling. The rest of this section is devoted to addressing this challenge.

Since $\widehat{\mathbf{B}}$ and $\widehat{\mathbf{V}}$ are jointly detected and estimated, correspondingly, if one is not reliable, neither is the other in general. We first examine the reliability of the bit matrix decision $\widehat{\mathbf{B}} = [\widehat{\mathbf{b}}_1, \ldots, \widehat{\mathbf{b}}_K]^T$ returned by the ILS procedure of Table I. The sample cross-correlation between any two bit streams is

$$\eta_{i,j} \triangleq \widehat{\mathbf{b}}_i^T \widehat{\mathbf{b}}_j / N,\ i \neq j,\ i, j = 1, \ldots, K. \tag{16}$$

Formally, the true information bits are independent within user streams and across users. If $\eta_{i,j}$ were to be viewed as approximately normally distributed with zero mean and variance $\frac{1}{N}$, then the probability of $|\eta_{i,j}|$, $i \neq j$, being larger than, say, the threshold value $\frac{3}{\sqrt{N}}$ is very low at about 0.3% (we can calculate $\Pr(|\eta_{i,j}| > \frac{3}{\sqrt{N}}) \approx 0.003$). Motivated by this calculation, we introduce below Criterion 1 that classifies convergence points of the ILS procedure in Table I as "unreliable" based on the sample statistics of the returned data matrix $\widehat{\mathbf{B}}$.

Criterion 1: If $|\eta_{i,j}| > \frac{3}{\sqrt{N}}$ for some $i \neq j \in \{1, 2, \ldots, K\}$, then ($\widehat{\mathbf{B}}$, $\widehat{\mathbf{V}}$) returned by the ILS procedure in Table I are classified as "*unreliable*." ■

Criterion 1 provides the means for coarse identification of unreliable solutions. An unreliable convergence point would then trigger re-initialization and re-execution of the ILS procedure in Table I. To enhance the end accuracy of the blind primary-system identification procedure, we propose one additional criterion based on the returned estimated effective signature matrix $\widehat{\mathbf{V}}$. We will motivate our proposal by examining the normalized cross-correlation between the estimated channel processed signatures $\widehat{\mathbf{v}}_i$ returned by the ILS procedure upon convergence and the true channel processed signatures $\mathbf{v}_i$, $i = 1, \ldots, K$. Based on $N = 128$ snapshots for a system with $K = 4$ primary users with equal 6dB SNR and processing gain $L = 31$ (Gold signature codes), we run the Criterion 1-equipped ILS procedure 20 times. The distribution of the twenty returned solutions of $\theta_i \triangleq \frac{\widehat{\mathbf{v}}_i^H \mathbf{v}_i}{\|\widehat{\mathbf{v}}_i\| \|\mathbf{v}_i\|}$ for signal $i = 1$ in Fig. 2 (representative of all other signals) reveals that (i) Criterion 1 is not by itself sufficient to eliminate erroneous solutions, however (ii) there exist "reliable" regions/clusters in which most of the Criterion 1-equipped ILS convergence points lie close to the true channel processed signatures. The basic idea behind our second and final refinement of the ILS blind primary-system identification procedure is to identify and average these reliable clustered estimates. Of course,

[4] In practical implementations, re-initialization and re-execution of the ILS procedure may also be needed whenever numerical instabilities create a rank deficient matrix $\widehat{\mathbf{B}}^{(d)}$ in Table I and $\widehat{\mathbf{B}}^{(d)}\widehat{\mathbf{B}}^{(d)T}$ is not invertible.

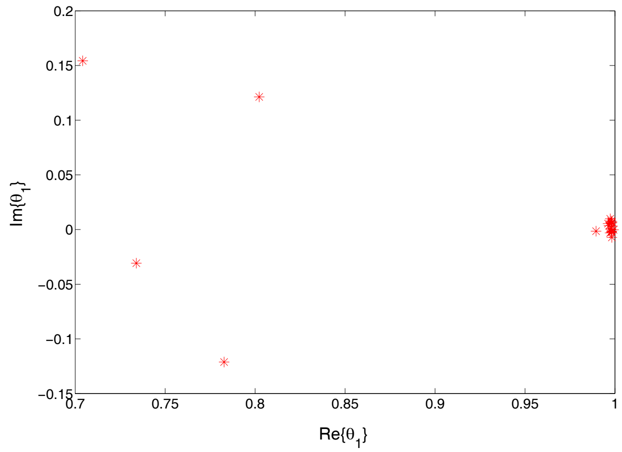

Fig. 2. Experimentation with ILS of Table I under Criterion 1: Distribution of normalized cross-correlation between $\widehat{\mathbf{v}}_1$ and $\mathbf{v}_1$ after twenty runs ($K = 4, L = 31, N = 128, \mathrm{SNR}_i = 6\mathrm{dB}, i = 1, 2, 3, 4$).

identification of the reliable estimates is not a trivial task due to our complete lack of knowledge of $\mathbf{v}_i$ (or $\mathbf{s}_i$).

In this context, assume that we have D estimates of $\mathbf{v}_i$ denoted by $\widehat{\mathbf{v}}_i^{(j)}$, $i = 1, \ldots, K$, $j = 1, \ldots, D$, obtained by D runs of the Criterion 1-equipped ILS procedure. From the example of Fig. 2, we notice that reliable estimates $\widehat{\mathbf{v}}_i^{(j)}$ of $\mathbf{v}_i$ have high normalized cross-correlation (close to 1) with each other, while they have low normalized cross-correlation with other unreliable estimates of $\mathbf{v}_i$. In contrast, unreliable estimates tend to have low normalized cross-correlation with each other. Therefore, the reliability of $\widehat{\mathbf{v}}_i^{(j)}$ may be quantified by examining the sum-cross-correlation with the rest $\widehat{\mathbf{v}}_i^{(t)}$, $t \neq j$,

$$\rho_i^{(j)} \triangleq \sum_{t=1, t\neq j}^{D} \frac{|\widehat{\mathbf{v}}_i^{(j)H} \widehat{\mathbf{v}}_i^{(t)}|}{\|\widehat{\mathbf{v}}_i^{(j)}\| \|\widehat{\mathbf{v}}_i^{(t)}\|}. \tag{17}$$

A reasonable threshold value for binary reliability classification may be the average value

$$\overline{\rho}_i \triangleq \frac{1}{D} \sum_{j=1}^{D} \rho_i^{(j)}, \; i = 1, \ldots, K, \tag{18}$$

utilized in the proposed Criterion 2 below.

Criterion 2: Let $\widehat{\mathbf{v}}_i^{(j)}$ be the estimates of $\mathbf{v}_i$ returned by D arbitrary initializations of the Criterion 1-equipped ILS procedure of Table I, $i = 1, \ldots, K$, $j = 1, \ldots, D$. If $\rho_i^{(j)} \geq \overline{\rho}_i$, then $\widehat{\mathbf{v}}_i^{(j)}$ is considered a *reliable* estimate of the $\mathbf{v}_i$; otherwise we declare it as *unreliable*. ■

Next, we average our reliable (according to *Criterion 2*) estimates of the channel processed signatures $\mathbf{v}_i$ to produce one last high-quality initialization of the ILS algorithm of Table I. Let $\mathcal{S}_i$ denote the set of all reliable estimates of $\mathbf{v}_i$ according to *Criterion 2* and let $|\mathcal{S}_i|$ denote the cardinality of $\mathcal{S}_i$. Our averaged estimate of the matrix $\mathbf{V}$ is now given by $\overline{\widehat{\mathbf{V}}}$ with

$$\overline{\widehat{\mathbf{V}}} \triangleq \left[\overline{\widehat{\mathbf{v}}}_1, \ldots, \overline{\widehat{\mathbf{v}}}_K\right] \text{ where } \overline{\widehat{\mathbf{v}}}_i = \frac{1}{|\mathcal{S}_i|} \sum_{j \in \mathcal{S}_i} \widehat{\mathbf{v}}_i^{(j)}, \; i = 1 \ldots, K, \tag{19}$$

i.e. $\overline{\widehat{\mathbf{v}}}_i$ is the average over all reliable estimates of $\mathbf{v}_i$ according to *Criterion 2*. We execute ILS in Table I a final time initialized at $\widehat{\mathbf{B}}^{(0)} = \mathrm{sgn}\left\{\mathfrak{Re}\left[\left(\overline{\widehat{\mathbf{V}}}^H \overline{\widehat{\mathbf{V}}}\right)^{-1} \overline{\widehat{\mathbf{V}}}^H \mathbf{Y}\right]\right\}$. We dub ILS with both Criteria 1 and 2 incorporated Cross-Correlation Enhanced ILS (CC-ILS) and summarize the complete procedure in Table II.

TABLE II
CROSS-CORRELATION ENHANCED ILS

For $j := 1$ to D
  1) Execute ILS of Table I with arbitrary initialization and obtain estimates $\widehat{\mathbf{v}}_i$, $i = 1, \ldots, K$.
  2) **If** estimates are *reliable* according to *Criterion 1*,
    let $\widehat{\mathbf{v}}_i^{(j)} := \widehat{\mathbf{v}}_i$, $i = 1, \ldots, K$;
    else go to 1).
End
For $i := 1$ to K
  3) Identify reliable estimates of $\mathbf{v}_i$ according to *Criterion 2*.
  4) Calculate average over all reliable estimates $\overline{\widehat{\mathbf{v}}}_i$ by (19).
End
5) Set $\overline{\widehat{\mathbf{V}}} \triangleq [\overline{\widehat{\mathbf{v}}}_1, \ldots, \overline{\widehat{\mathbf{v}}}_K]$.
6) Execute ILS of Table I with initialization
$$\widehat{\mathbf{B}}^{(0)} = \mathrm{sgn}\left\{\mathfrak{Re}\left[\left(\overline{\widehat{\mathbf{V}}}^H \overline{\widehat{\mathbf{V}}}\right)^{-1} \overline{\widehat{\mathbf{V}}}^H \mathbf{Y}\right]\right\}.$$

B. Detection of Primary Binary Signatures

After obtaining an estimated (energy inclusive) channel processed signature set $\widehat{\mathbf{V}}$ by CC-ILS of Table II, we develop another procedure to extract the individual primary binary signatures $\mathbf{s}_i$, $i = 1, \ldots, K$, by decomposition of $\widehat{\mathbf{V}}$. The channel processed signatures can be rewritten as

$$\mathbf{v}_i = \sqrt{E_i}\mathbf{S}_i\mathbf{h}, \; i = 1, \ldots, K, \tag{20}$$

where $\mathbf{h} = [h_1, \ldots, h_M]^T$ and

$$\mathbf{S}_i \triangleq \begin{bmatrix} s_i(1) & & \mathbf{0} \\ & \ddots & \\ \vdots & & s_i(1) \\ s_i(L) & & \vdots \\ & \ddots & \\ \mathbf{0} & & s_i(L) \end{bmatrix}_{L_M \times M}. \tag{21}$$

If the binary signatures $\mathbf{s}_i$, $i = 1, \ldots, K$, were known, by (20) we could estimate $\sqrt{E_i}\mathbf{h}$ as

$$\widehat{\sqrt{E_i}\mathbf{h}} = (\mathbf{S}_i^T\mathbf{S}_i)^{-1}\mathbf{S}_i^T\widehat{\mathbf{v}}_i, \; i = 1, \ldots, K, \tag{22}$$

where $\widehat{\mathbf{v}}_i$ is the ith column of matrix $\widehat{\mathbf{V}}$. Then, a quality estimate of $\mathbf{h}$ (scaled) could be produced by averaging,

$$\widehat{\mathbf{h}} = \frac{1}{K} \sum_{i=1}^{K} \widehat{\sqrt{E_i}\mathbf{h}}, \tag{23}$$

to create a matrix channel estimate $\widehat{\mathbf{H}}$ by (2). Given $\widehat{\mathbf{H}}$ (and $\widehat{\mathbf{v}}_i$), one could detect the binary signatures by

$$\widehat{\mathbf{s}}_i = \frac{1}{\sqrt{L}}\mathrm{sgn}\left\{\mathfrak{Re}\{(\widehat{\mathbf{H}}^H\widehat{\mathbf{H}})^{-1}\widehat{\mathbf{H}}^H\widehat{\mathbf{v}}_i\}\right\}, i = 1, \ldots, K. \tag{24}$$

The proposed individual binary signature extraction algorithm from $\widehat{\mathbf{V}}$ is now ready. Initialize $\mathbf{s}_i \in \frac{1}{\sqrt{L}}\{\pm1\}^L$, $i = 1, \ldots, K$,

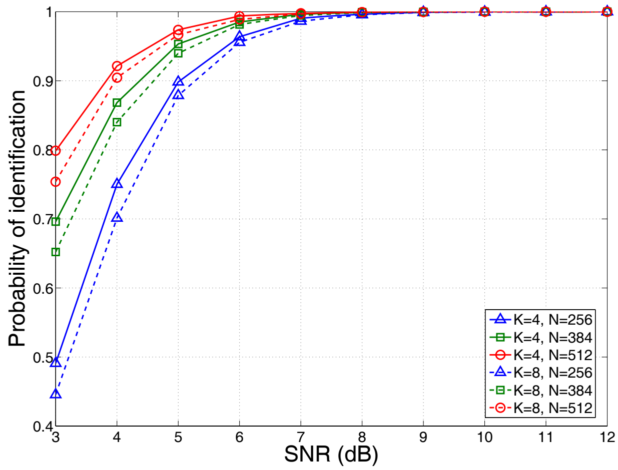

Fig. 3. Performance of blind primary-users identification algorithm: Primary transmitter PT has K downlink users with length $L = 31$ Gold signatures. The transmitted signal propagates over a 3-path Rayleigh fading channel to secondary receiver SR that collects N observation samples to run Table II procedure followed by execution of (22)-(24).

arbitrarily, and alternate computation between (22), (23), and (24), iteratively. Stop when convergence is observed[5].

To illustrate briefly the proposed primary binary signature detection algorithm, we consider a primary downlink CDMA system with $K = 4$ or $K = 8$ active users that utilize Gold signatures with length $L = 31$. The primary users' signals are transmitted with equal power over a multipath Rayleigh fading channel with $M = 3$ resolvable paths in the presence of additive white Gaussian noise. SR is able to collect $N = 256, 384$, or 512 observation samples and employs our proposed method to extract the primary users' binary signatures. The experiment is repeated 10^5 times with randomly drawn channel coefficients. In Fig. 3, we plot the probability of correct identification of *all* binary signatures, $\Pr(\widehat{\mathbf{s}}_i = \mathbf{s}_i \, \forall i = 1, \ldots, K)$, as a function of SNR. It can be seen that even under low/moderate SNR values the proposed method can correctly extract all signatures with sufficient sample support (secondary receiver observation time interval). It is worth pointing out that, experimentally, when errors do occur only one or two signatures have few chip-bit errors only.

Having addressed the blind primary system identification problem -arguably- satisfactorily, in the next section we attempt to solve the secondary power and code-channel allocation problem of (10) to enable most frequent activation of the secondary link.

IV. Secondary Code-Division Link Channelization

The secondary link has to satisfy the SINR constraint (11); namely, that the SINR of the secondary receiver SR at the maximum-SINR linear filter output is no less than a QoS requirement γ: $P\mathbf{c}^T\mathbf{G}^H\mathbf{R}\mathbf{G}\mathbf{c} \geq \gamma$, $\gamma > 0$. To that respect, it suffices to set the power value $P > 0$ to

$$P = \frac{\gamma}{\mathbf{c}^T\mathbf{G}^H\mathbf{R}\mathbf{G}\mathbf{c}} \tag{25}$$

[5]The iterative procedure may be re-executed with distinct initialization if convergence cannot be observed after sufficient iterations.

and the SINR constraint (11) is always satisfied with equality. Then, the optimization problem in (10) is equivalent to

$$\underset{\mathbf{c}\in\frac{1}{\sqrt{L}}\{\pm 1\}^L}{\text{minimize}} \;\; \mathcal{I}_{max} = max\{P \cdot \text{PTSC}(\mathbf{c}, \mathbf{s}_i) : i = 1, \ldots, K\}$$
$$\text{s. t.} \;\; P = \frac{\gamma}{\mathbf{c}^T\mathbf{G}^H\mathbf{R}\mathbf{G}\mathbf{c}}. \tag{26}$$

The maximum generalized correlation interference $\mathcal{I}_{max}$ is the product of two components, transmit power P and maximum PTSC value. Thus, to minimize $\mathcal{I}_{max}$ we need to design a binary signature $\mathbf{c}$ to minimize the product of the required transmit power $P = \frac{\gamma}{\mathbf{c}^T\mathbf{G}^H\mathbf{R}\mathbf{G}\mathbf{c}}$ times $\underset{i=1,\ldots,K}{max}\{\text{PTSC}(\mathbf{c}, \mathbf{s}_i)\}$. This problem is still non-convex NP-hard (see Footnote 1 and [12], [13]). We must, therefore, pursue (disjoint) suboptimal design procedures if we wish to keep the computational complexity manageable. At first, we look at minimizing P *alone*. The binary signature $\mathbf{c}$ that minimizes P maximizes the denominator of (25) which is the SR output SINR with unit ST transmit power:

$$\mathbf{c} = \arg \max_{\mathbf{c}\in\frac{1}{\sqrt{L}}\{\pm 1\}^L} \mathbf{c}^T\mathbf{A}\mathbf{c} \tag{27}$$

where $\mathbf{A} \triangleq \mathbf{G}^H\mathbf{R}\mathbf{G}$. At this point, the SINR-maximizing binary signature designs of polynomial complexity developed in [26], [27] can be used directly. We recall that in [26] the binary signature vector is optimized under a rank-2 approximation of the matrix $\mathbf{A}$, while in [27] the arcs of least SINR decrease from the real maximum SINR solution are evaluated. Both algorithms first generate L candidate binary signatures[6], denoted by $\mathbf{q}_j \in \frac{1}{\sqrt{L}}\{\pm 1\}^L$, $j = 1, \ldots, L$, which can provide high output SINR. Then, the signature among them with highest output SINR is selected. While the highest-SINR signature minimizes the transmitting power P required to satisfy any given QoS constraint, it may result to high values of PTSC with respect to primary user signatures and consequently let the secondary link introduce strong interference to the primary system. Therefore, we propose to evaluate all L binary signatures $\mathbf{q}_j \in \frac{1}{\sqrt{L}}\{\pm 1\}^L$, $j = 1, \ldots, L$, returned by the solver of (27) in [26] or [27] in our interference metric for the ith primary user

$$\mathcal{I}_{j,i} = P_j \cdot \text{PTSC}(\mathbf{q}_j, \mathbf{s}_i), \; j = 1, \ldots, L; \; i = 1, \ldots, K, \tag{28}$$

and find the maximal generalized correlation interference value caused by $\mathbf{q}_j$, $\mathcal{I}_{j,max} = \max\{\mathcal{I}_{j,i} : i = 1, \ldots, K\}$, $j = 1, \ldots, L$. Then, we choose the signature, power pair $(\mathbf{q}_{j^*}, P_{j^*})$ which has the least maximal interference. If the resulting maximal interference by $(\mathbf{q}_{j^*}, P_{j^*})$ is introduced to the i^*th primary user, i.e. $\mathcal{I}_{j^*,i^*} = \min\{\mathcal{I}_{j,max} : j = 1, \ldots, L\}$, and still $\mathcal{I}_{j^*,i^*} < \mathcal{I}_{\text{th}}$, the secondary link is allowed to access the channel by assigning signature $\mathbf{c} = \mathbf{q}_{j^*}$ and power $P = P_{j^*}$. We refer to this method of selecting a pair of signature and power from the candidate set as *passive interference suppression* and outline the procedure in Table III.

The secondary link design method in Table III exclusively focuses on transmit power minimization. The PTSC factor

[6]We recall that [27] finds $(LT - T + 1)$ binary sequences that are closest to T arcs of least SINR decrease in the l_2 sense. In this paper, we only consider one slowest descent arc and generate L binary sequences. This is sufficient to closely approximate the performance level reached when all $L - 1$ slowest descent arcs are considered.

TABLE III
SECONDARY LINK SIGNATURE AND POWER DESIGN W/ PASSIVE INTERFERENCE SUPPRESSION

Input $\mathbf{A} := \mathbf{G}^H\mathbf{R}\mathbf{G}$
Obtain $\mathbf{q}_j$, $j = 1, \ldots, L$ as solution candidates for (27) by [26] or [27].
Calculate P_j, $j = 1, \ldots, L$, by (25).
Calculate $\mathcal{I}_{j,i}$, $j = 1, \ldots, L$, $i = 1, \ldots, K$, by (28) and
$\mathcal{I}_{j,max} = max\{\mathcal{I}_{j,i} : i = 1, \ldots, K\}$.
Select j^*, i^* such that $\mathcal{I}_{j^*,i^*} = min\{\mathcal{I}_{j,max} : j = 1, \ldots, L\}$.
Output P_{j^*}, $\mathbf{q}_{j^*}$, and $\mathcal{I}_{j^*,i^*}$.
If $\mathcal{I}_{j^*,i^*} < \mathcal{I}_{\text{th}}$,
transmit on channel q_{j^*} with power P_{j^*};
else seize

TABLE IV
SECONDARY LINK SIGNATURE AND POWER DESIGN W/ ACTIVE INTERFERENCE AVOIDANCE

Input $\mathbf{R}$, $\mathbf{G}$, and $\mathbf{s}_i$, $i = 1, \ldots, K$
$d := 0$.
$\mathbf{A} := \mathbf{G}^H\mathbf{R}\mathbf{G}$.
Obtain P_{j^*}, $\mathbf{q}_{j^*}$, and $\mathcal{I}_{j^*,i^*}$ by Table III.
$P^d := P_{j^*}$, $\mathbf{c}^d := \mathbf{q}_{j^*}$, $\mathcal{I}^d_{max} := \mathcal{I}_{j^*,i^*}$.
While $d = 0$ or $\mathcal{I}^d_{max} < \mathcal{I}^{(d-1)}_{max}$
$d := d + 1$;
$\mathbf{A} \leftarrow \mathbf{A} - \alpha\widetilde{\mathbf{S}}_{i^*}\widetilde{\mathbf{S}}^T_{i^*}$;
obtain P_{j^*}, $\mathbf{q}_{j^*}$, and $\mathcal{I}_{j^*,i^*}$ by Table III;
$P^d := P_{j^*}$, $\mathbf{c}^d := \mathbf{q}_{j^*}$, $\mathcal{I}^d_{max} := \mathcal{I}_{j^*,i^*}$.
End
Output $P^{(d-1)}$, $\mathbf{c}^{(d-1)}$, and $\mathcal{I}^{(d-1)}_{max}$.
If $\mathcal{I}^{d-1}_{max} < \mathcal{I}_{\text{th}}$,
transmit on channel $\mathbf{c}^{(d-1)}$ with power $P^{(d-1)}$;
else seize.

is, of course, evaluated and accounted for in (28) but not actively optimized (minimized). To further reduce the maximum interference caused to the primary system and further improve the chances of spectrum sharing, we next propose to iteratively adjust the binary channel signature to actively avoid interference by jointly reducing the PTSC value with the most impacted at each time i^*th primary user's signature $\mathbf{s}_{i^*}$, as well as minimize the transmit power.

Define $\widetilde{\mathbf{S}}_{i^*} \triangleq \left[\mathbf{s}_{i^*|-M}, \ldots, \mathbf{s}_{i^*|0}, \ldots, \mathbf{s}_{i^*|M}\right]$ and calculate $\text{PTSC}(\mathbf{c}, \mathbf{s}_{i^*}) = \mathbf{c}^T\widetilde{\mathbf{S}}_{i^*}\widetilde{\mathbf{S}}^T_{i^*}\mathbf{c}$. To combine the SINR-optimization problem of (27) with the PTSC suppression task, after executing the procedure in Table III we update $\mathbf{A} = \mathbf{G}^H\mathbf{R}\mathbf{G} - \alpha\widetilde{\mathbf{S}}_{i^*}\widetilde{\mathbf{S}}^T_{i^*}$ where $\alpha > 0$ is an introduced weighting factor. Then, by re-executing the procedure in Table III with the updated $\mathbf{A}$, we obtain a new optimal pair $(\mathbf{q}_{j^*}, P_{j^*})$ and new maximum-interference $\mathcal{I}_{j^*,i^*}$. If the new maximum-interference $\mathcal{I}_{j^*,i^*}$ is reduced, we iteratively update $\mathbf{A} \leftarrow \mathbf{A} - \alpha\widetilde{\mathbf{S}}_{i^*}\widetilde{\mathbf{S}}^T_{i^*}$ and re-execute the algorithm in Table III. This procedure will be stopped when the maximum interference cannot be suppressed further. We outline the procedure in Table IV.

Duplexing of the secondary link can be implemented in TDD mode by switching the roles of transmitter and receiver. The reverse secondary link can be activated by a new pair of transmit power and channel code/signature values calculated by the same algorithm in Table III or IV to satisfy reverse link QoS requirements. The following section is devoted to the generalization of the channelization problem to multiple secondary receivers (one-to-many secondary downlink transmissions).

V. DESIGN OF MULTIPLE SECONDARY CODE-DIVISION LINKS

In this section, we extend the study to cover code-division channelization for multiple secondary links. Particularly, as shown in Fig. 4, we consider the scenario where an ST attempts to communicate downlink with Q potential SRs[7]. Let $\mathcal{C} = \{1, \ldots, Q\}$ be the set of all secondary link indices which attempt to share the radio spectrum. Among the Q potential secondary links, $K_S \geq 0$ secondary links will be activated by assigning to each a (normalized) signature $\mathbf{c}_k = \frac{1}{\sqrt{L}}\{\pm1\}^L$ and transmitting power $P_k > 0$, $k \in \mathcal{S} \subseteq \mathcal{C}$ where $\mathcal{S}$ represents the set of active secondary links.

[7]Multiple one-to-many secondary downlinks may be treated sequentially establishing one secondary downlink at a time using techniques presented in this section, treating previously established secondary downlinks as part of a "virtual primary system."

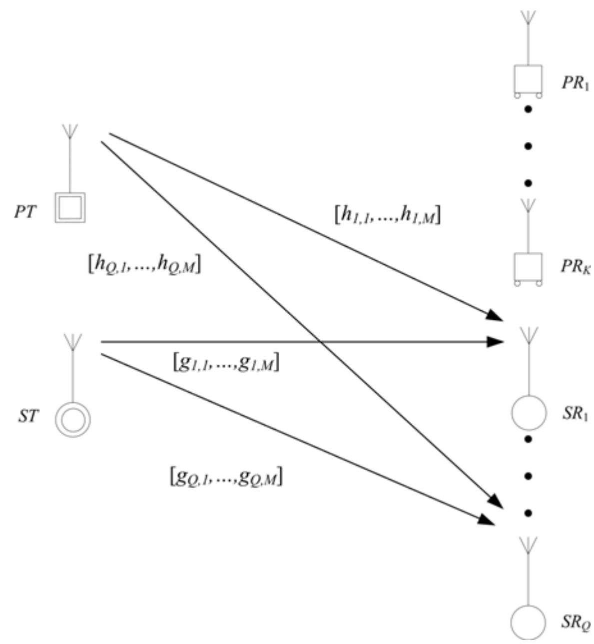

Fig. 4. Primary/secondary code-division system model of a primary transmitter PT, K primary receivers $PR_i, i = 1, 2, \ldots, K$, a secondary transmitter ST, and Q secondary receivers $SR_i, i = 1, 2, \ldots, Q$ (all receives signals exhibit multipath Rayleigh fading).

The discrete time received signal vector of the kth SR can be expressed as

$$\mathbf{r}_k(n) = \sqrt{P_k}b_k(n)\mathbf{G}_k\mathbf{c}_k + \sum_{j\in\mathcal{S}, j\neq k}\sqrt{P_j}b_j(n)\mathbf{G}_k\mathbf{c}_j + \mathbf{y}_k(n),$$
$$k \in \mathcal{S}, n = 1, 2, \ldots, \tag{29}$$

where $\mathbf{G}_k \in \mathbb{C}^{L_M\times L}$ is the channel matrix constructed by the multipath channel coefficients $[g_{k,M}, \ldots, g_{k,M}]$ between the secondary transmitter and the kth secondary receiver and $\mathbf{y}_k$ represents comprehensively MAI from all primary users plus AWGN. The autocorrelation matrix of $\mathbf{y}_k$ is denoted by $\mathbf{R}_k \triangleq \mathrm{E}\left\{\mathbf{y}_k\mathbf{y}^H_k\right\}$. The output SINR of the kth secondary receiver filter $\mathbf{w}_{maxSINR,k}$ is given by

$$\Gamma_k = P_k\mathbf{c}^T_k\mathbf{G}^H_k\widetilde{\mathbf{R}}^{-1}_k\mathbf{G}_k\mathbf{c}_k \tag{30}$$

where

$$\widetilde{\mathbf{R}}_k = \mathbf{G}_k \left(\sum_{j \in \mathcal{S}, j \neq k} P_j \mathbf{c}_j \mathbf{c}_j^T \right) \mathbf{G}_k^H + \mathbf{R}_k \tag{31}$$

is the autocorrelation matrix of the combined channel disturbance. To satisfy a certain QoS requirement, each active secondary link is supposed to surpass a pre-determined SINR value $\gamma_k > 0$, i.e. we need $\Gamma_k \geq \gamma_k$ for each $k \in \mathcal{S}$. At the same time, interference caused to the primary system needs to be kept in check. To evaluate the interference introduced by all active secondary links to the ith primary user, we define the *cumulative generalized correlation interference*

$$\mathcal{I}_i = \sum_{k \in \mathcal{S}} P_k \cdot \mathrm{PTSC}(\mathbf{c}_k, \mathbf{s}_i), \quad i = 1, \ldots, K. \tag{32}$$

Our objective is to maximize the number of active secondary links $K_S = |\mathcal{S}|$ under the constraints that each secondary link achieves its pre-determined SINR γ_k and the cumulative interference to each primary user is less than a threshold $\mathcal{I}_{\text{th}} > 0$:

$$\begin{gathered} \text{maximize } K_S \\ \text{s. t. } \Gamma_k(\mathbf{c}_k, P_k) \geq \gamma \, \forall \, k \in \mathcal{S} \text{ and } \mathcal{I}_i < \mathcal{I}_{\text{th}} \; \forall \, i = 1, \ldots, K. \end{gathered} \tag{33}$$

This problem is, again, non-convex NP-hard. Motivated by the single secondary link channelization algorithm of Table IV (with active interference avoidance) in the previous section, we propose instead to -centrally at the secondary transmitter (base station) ST- successively activate one secondary link at a time in a way that it introduces conditionally least cumulative interference until the accumulated interference exceeds the threshold $\mathcal{I}_{\text{th}}$ for some primary user. The details of the algorithm are described as follows. Initially set $\mathcal{S} = \varnothing$. Utilizing the algorithm in Table IV, we individually evaluate each secondary link $k \in \mathcal{C}$ and obtain the signature $\mathbf{c}_k$, transmitting power P_k, and the maximum generalized interference value $\mathcal{I}_{max,k}$ caused to a primary user by the potential secondary link with design $(\mathbf{c}_k, P_k)$. Among the Q potential links, we select the one, say the tth secondary link, which introduces least maximum generalized interference, i.e. $\mathcal{I}_{max,t} = min\{\mathcal{I}_{max,k} : k \in \mathcal{C}\}$. If $\mathcal{I}_{max,t} < \mathcal{I}_{\text{th}}$, the tth secondary link is turned on with code channel $\mathbf{c}_t$ and power P_t; if link t did go through, we continue on examining the remaining $\mathcal{C} - \{t\}$ candidates by the algorithm in Table IV under the cumulative generalized interference metric in (32) since one secondary link is already activated. Say potential secondary link $z \in \mathcal{C} - \{t\}$ is the one with $\mathcal{I}_{max,z} = min\{\mathcal{I}_{max,k} : k \in \mathcal{C} - \{t\}\}$ and still $\mathcal{I}_{max,z} < \mathcal{I}_{\text{th}}$ (if $\mathcal{I}_{max,z} \geq \mathcal{I}_{\text{th}}$, link z is rejected and no further study is required). Since the incoming secondary link z also interferes with the existing secondary link t and may degrade its output SINR below the minimum acceptable level, we adjust iteratively the transmitting power of each secondary link by calculating

$$P_k = \frac{\gamma_k}{\mathbf{c}_k^T \mathbf{G}_k^H \widetilde{\mathbf{R}}_k^{-1} \mathbf{G}_k \mathbf{c}_k}, \; k = t, z, \tag{34}$$

until convergence is observed. With the updated powers, we evaluate the cumulative generalized interference in (32) one more time. If all $\mathcal{I}_i, i = 1, \ldots, K$, are below the threshold, the zth secondary link is turned on and we move on to identify additional secondary links with the same procedure; otherwise, z is rejected and no more secondary link are allowed to share the spectrum. The complete algorithm is summarized in Table V.

TABLE V
COGNITIVE CODE-DIVISION CHANNELIZATION FOR MULTIPLE SECONDARY LINKS

$\mathcal{C} := \{1, \ldots, Q\}$; $\mathcal{S} := \varnothing$.
While $|\mathcal{C}| > 0$
  Calculate $\widetilde{\mathbf{R}}_k \; \forall k \in \mathcal{C}$ by (31).
  Obtain P_k, $\mathbf{c}_k$ $\mathcal{I}_{max,k}$ $\forall k \in \mathcal{C}$ by Table IV.
  Find $t \in \mathcal{C}$ s. t. $\mathcal{I}_{max,t} = min\{\mathcal{I}_{max,k} : k \in \mathcal{C}\}$.
  Update powers P_k, $k \in \{\mathcal{S}, t\}$ by (34).
  Calculate $\mathcal{I}_i$, $i = 1, \ldots, K$, by (32) (with the updated powers).
  If $\mathcal{I}_i < \mathcal{I}_{\text{th}} \; \forall \, i = 1, \ldots, K$,
    $\mathcal{C} = \mathcal{C} \backslash \{t\}$; $\mathcal{S} = \mathcal{S} \bigcup \{t\}$.
  else
    $\mathcal{C} = \varnothing$.
  Endif
End
Output $\{(P_j, \mathbf{c}_j) : j \in \mathcal{S}\}$ (activated secondary channels).

VI. EXPERIMENTAL SIMULATION STUDIES

To illustrate the presented algorithmic developments, we consider a primary multiuser CDMA system with K active users that utilize Gold signatures with length $L = 31$. The primary users' signals are transmitted with equal per-user power $E_1 = E_2 = \ldots = E_K = 10$dB over a multipath Rayleigh fading channel with $M = 3$ resolvable paths in the presence of additive white Gaussian noise. At first, one secondary link attempts to share the spectrum with target receiver filter output SINR $\gamma = 10$dB.

The secondary link code channel and power are optimized by the algorithm in Table III (passive interference suppression) and Table IV (active interference avoidance). Specifically, we utilize the rank-2 SINR-maximization binary signature design method [26] and set the weighting factor for the algorithm in Table IV to $\alpha = 0.05$. If the resulting maximum generalized correlation interference is less than the threshold $\mathcal{I}_{\text{th}}$, the secondary link is allowed to share the spectrum. The simulation experiment is repeated 10^5 times. With varying primary user population $K = 2, 10, 18$, the probability of coexistence of a secondary link is plotted in Fig. 5 as a function of $\mathcal{I}_{\text{th}}$. While the Table III algorithm behaves -arguably- satisfactorily, it can be observed that the active interference avoidance algorithm of Table IV can significantly enhance the opportunity of coexistence of a secondary link.

Next, under the same primary system simulation environment, we examine the problem of cognitive code-division channelization for multiple secondary links. The receiver output SINR requirement for each secondary link is set at $\gamma_j = 10$dB, $j = 1, 2, \ldots$. The average number of activated secondary links by the algorithm in Table V is plotted in Fig. 6 as a function of the interference threshold $\mathcal{I}_{\text{th}}$ for varying primary user population K. In Fig. 7, we plot the average number of active secondary links versus the number of primary users for varying interference threshold values $\mathcal{I}_{\text{th}}$. The experiments demonstrate that the developed multiple secondary link channelization method can effectively produce multiple active secondary links and enhance spectrum efficiency.

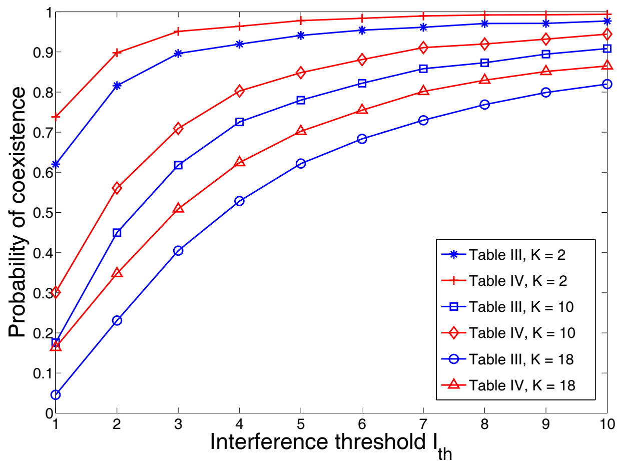

Fig. 5. Probability of coexistence of secondary link versus interference threshold $\mathcal{I}_{\mathrm{th}}$.

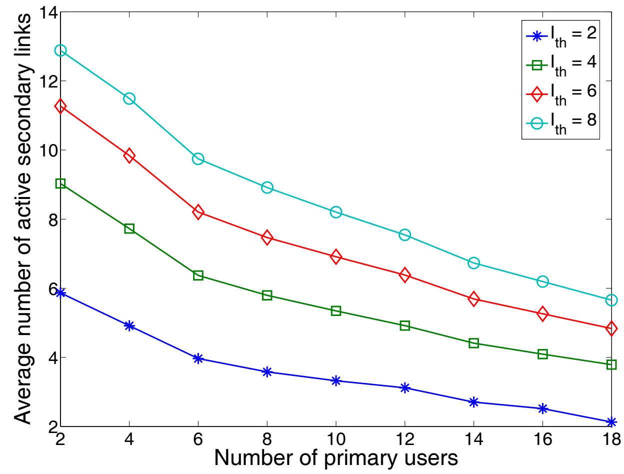

Fig. 7. Average number of secondary links versus number of primary users K.

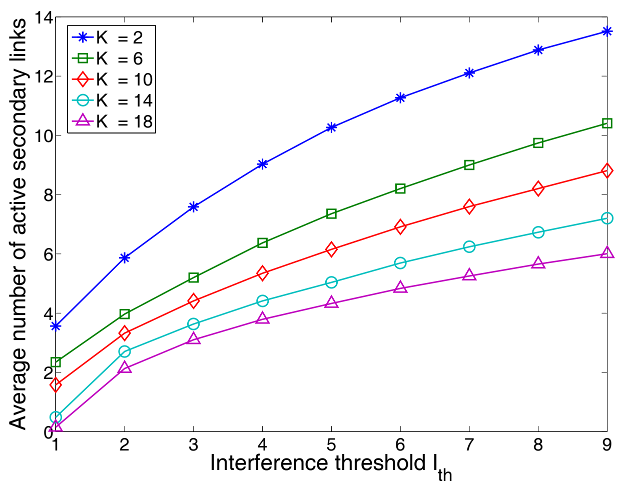

Fig. 6. Average number of activated secondary links by the algorithm in Table V versus interference threshold $\mathcal{I}_{\mathrm{th}}$ for varying primary user population K.

VII. Conclusions

We considered the general problem of establishing a secondary code-division link alongside a primary code-division multiple-access system. We first developed a novel iterative least square (ILS) based primary-system identification algorithm which can blindly detect the code channels utilized by primary users. Then, we proposed two alternative schemes, one of low (passive scheme) and one of moderate (active scheme) computational complexity that optimize transmitting power and binary code-channel allocation of the secondary link without causing "harmful" interference to the primary users. At the same time, the signal-to-interference-plus-noise ratio (SINR) of the secondary link at the output of the maximum SINR linear receiver is no less than a certain threshold to meet quality of service (QoS) requirements for the secondary link. Finally, we extended the channelization problem to multiple code-division secondary links.

Simulation results demonstrated that the proposed blind identification algorithm can efficiently and effectively detect primary users' code channels and the proposed code-division channelization methods can successfully allow secondary links to opportunistically share the spectrum without causing harmful interference to primary users.

Cognitive code-division radios combine in principle the bandwidth efficiency characteristics of cognitive operation and code-division multiple accessing and are expected to find a place in future communication systems. To that extend, the developments presented in this paper constitute an early contribution that can be helpful in benchmarking future efforts. Since the developed herein coexistence-mode cognitive networks do not require any prior knowledge about the primary networks (such as signatures, transmission energy, channel state information), the secondary networks can be deployed transparently without any modification/upgrade of the existing primary CDMA infrastructure.

Appendix A
Derivation of Iterative Least-Squares Procedure in Table I

The LS cost function in (15) can be rewritten as

$$J = \|\mathbf{Y}-\mathbf{V}\mathbf{B}\|_F^2 = \mathrm{tr}\left\{\mathbf{Y}\mathbf{Y}^H\right\} - \mathrm{tr}\left\{\mathbf{Y}\mathbf{B}^H\mathbf{V}^H\right\} - \mathrm{tr}\left\{\mathbf{V}\mathbf{B}\mathbf{Y}^H\right\} + \mathrm{tr}\left\{\mathbf{V}\mathbf{B}\mathbf{B}^H\mathbf{V}^H\right\} \quad (35)$$

where $\mathrm{tr}\{\cdot\}$ denotes the trace of a matrix.

For a given $\mathbf{B}$, the LS optimal estimate of $\mathbf{V}$ can be obtain by differentiating the cost function J with respect to $\mathbf{V}^H$ and setting the outcome equal to the zero matrix,

$$\frac{\partial J}{\partial \mathbf{V}^H} = -\mathbf{Y}\mathbf{B}^H + \mathbf{V}(\mathbf{B}\mathbf{B}^H) = \mathbf{0} \Rightarrow \mathbf{V} = \mathbf{Y}\mathbf{B}^H(\mathbf{B}\mathbf{B}^H)^{-1}. \quad (36)$$

Next, pretend that $\mathbf{V}$ is known and relax the domain of the symbol information matrix to the complex space, $\mathbf{B} \in \mathbb{C}^{K\times N}$. The LS optimal estimate of $\mathbf{B} \in \mathbb{C}^{K\times N}$ can be calculated again by differentiation

$$\frac{\partial J}{\partial \mathbf{B}^H} = -\mathbf{V}^H\mathbf{Y} + \mathbf{V}^H\mathbf{V}\mathbf{B} = \mathbf{0} \Rightarrow \mathbf{B} = (\mathbf{V}^H\mathbf{V})^{-1}\mathbf{V}^H\mathbf{Y}. \quad (37)$$

Finally, we project (quantize) the complex-valued LS estimate $\mathbf{B}$ to the binary domain

$$\mathbf{B} = \mathrm{sgn}\left\{\mathfrak{Re}\left[(\mathbf{V}^H\mathbf{V})^{-1}\mathbf{V}^H\mathbf{Y}\right]\right\}. \quad (38)$$

■

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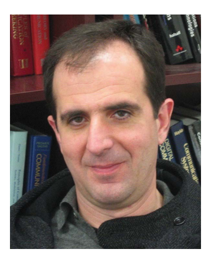

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